Using a portable-power boost converter in an isolated flyback application

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Some electronic applications require an isolated auxiliary power supply that must be generated from a low-voltage AC adapter. Examples of such applications include isolated MOSFET drive circuits and relay controls. Finding an isolated power-supply controller that will operate at such low voltages can prove difficult. Fortunately, it is possible to configure some low-power boost converters, such as the Texas Instruments (TI) TPS61175, to operate and control a flyback power stage. This article explains how to use the TPS61175 in an isolated flyback application.

Consider, for example, an application that is supplied from a 5-V wall wart and requires an isolated 5-V, 500-mA auxiliary supply. In this application, the input voltage can range from 3 to 6 V, and small size is a higher priority than efficiency. The low output-power and isolation requirements make the flyback an obvious choice for the power-supply topology. Taking into account the size limitations, a boost controller with an integrated FET is ideal for this type of application. With its 2.9-V minimum input voltage, the TPS61175 is one of the few boost converters that can meet these input-voltage specifications. In addition, the switching frequency of up to 2.2 MHz allows for a smaller transformer and reduced component sizes for the input and output filters.

The TPS61175 is a highly integrated current-mode controller. This controller internally limits the peak FET current to 3 A, which makes continuous conduction mode (CCM) a good choice. Selecting an operating frequency of 1 MHz keeps the switching losses manageable and reduces the transformer inductance and physical size. Although the TPS61175 includes a fixed amount of internal slope compensation, limiting the maximum duty cycle to less than 50% eliminates the possibility of bimodal operation and also reduces the rms current in the output capacitors.

Equation 1 computes the required primary-to-secondary turns ratio, N_{p2s}, of the transformer:

$$N_{p2s} = \frac{V_{IN(min)} \times D_{max}}{(V_{OUT} + V_d) \times (1 - D_{max})} \qquad (1)$$

In order to limit the duty cycle to 45%, a ratio of 1:2.5 was selected. The internal MOSFET of the TPS61175 is rated for 40 V, so the voltage stress is not an issue for this design.

In CCM converters, using too large an inductance can cause the right-half-plane zero (RHPZ) to limit the bandwidth of the feedback loop. The location of the RHPZ is determined by Equation 2:

$$RHPZ = R_{Load} \times \frac{(1-D)^2}{2\pi \times L_{Sec} \times D} \qquad (2)$$

If this frequency is not significantly higher than the unity-gain frequency of the feedback loop, stability is jeopardized. Fortunately, with a 1-MHz switching frequency, a relatively large inductance can be used and still allow for an adequate loop bandwidth. Selecting a primary inductance of 1.2 µH keeps the converter in CCM operation down to a load current of 150 mA and puts the RHPZ in the neighborhood of 200 kHz.

The amount of output capacitance required for a given peak-to-peak output ripple voltage is determined by Equation 3:

$$C_{min} = \frac{I_{OUT} \times \left\{1 - D \times \left[V_{IN(max)}\right]\right\}}{f_{SW} \times V_{PP}} \qquad (3)$$

In this design, using a single 22-µF ceramic capacitor limits the ripple voltage to 17 mV_{PP} with a 6-V input. Using another 22-µF ceramic capacitor on the input limits the input ripple voltage to similar levels.

Figure 1. TPS61175 boost controller used in simple isolated flyback

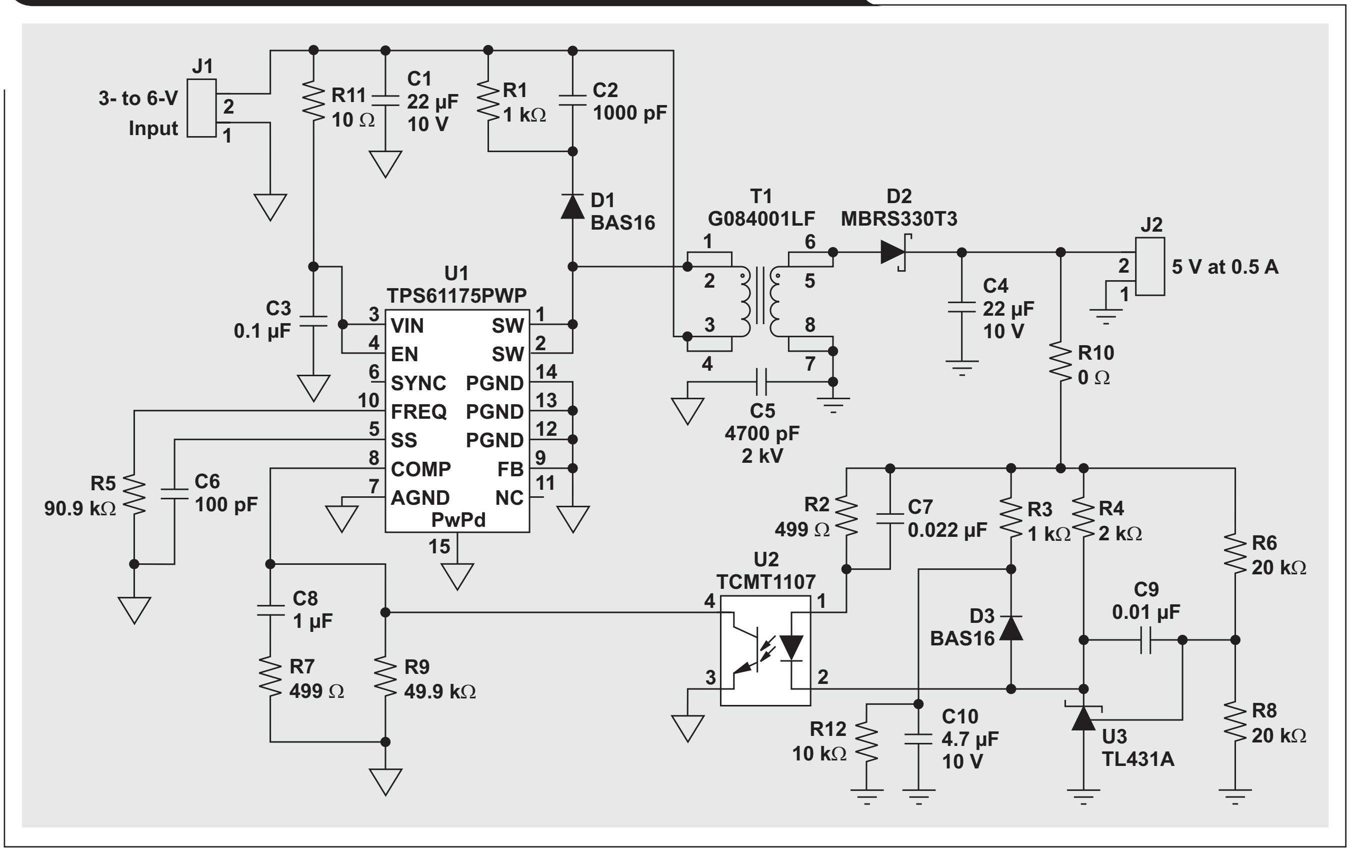


Figure 1 shows a schematic for this power-supply design. The TPS61175 contains a transconductance error amplifier. Normally, in nonisolated applications, the output voltage is fed back through a resistor divider to the feedback pin (FB, or pin 9). For an isolated feedback, the feedback pin must be grounded. This turns the output of the transconductance amplifier, located at the COMP pin, into a 130-µA current source. Connecting a 49.9-kΩ resistor (R9) from the COMP pin to ground allows the optocoupler to control the voltage of the COMP pin over its entire dynamic range (0.75 to 3 V).

The value of R2 must be small enough to allow the TI TL431 to sufficiently drive the optocoupler. The DC gain of the optocoupler circuit is determined by the current-transfer ratio (CTR), R2, and R9. Because the optocoupler needs to drive only 130 µA, the CTR is relatively low (approximately 10%), but the 100:1 ratio of R9 to R2 gives the optocoupler circuit a total DC gain of 20 dB. C8 and R7 were added to attenuate the gain and allow the loop to cross near 10 kHz. C8 and R9 form a pole at 30 Hz, while C8 and R7 form a zero at 3 kHz. This lowers the gain by 40 dB at frequencies above 3 kHz.

Figure 2. The optocoupler limits the bandwidth to around 10 kHz

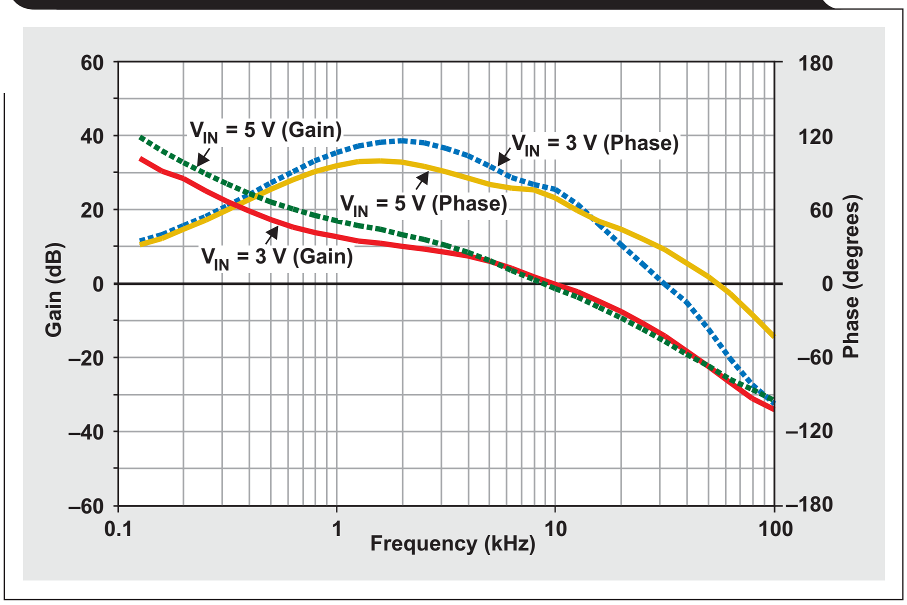
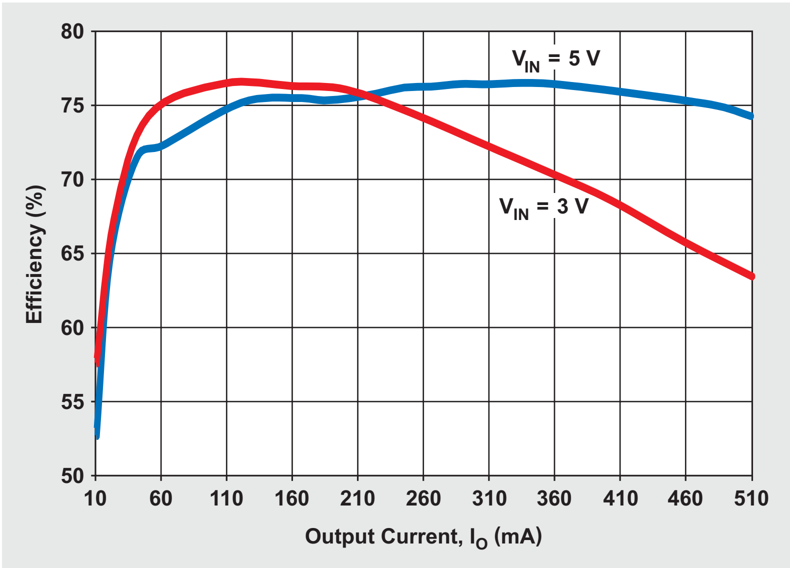


Figure 2 shows Bode plots of the feedback loop at full rated load for input voltages of 3 and 5 V. The response

Figure 3. Efficiency with a 5-V input



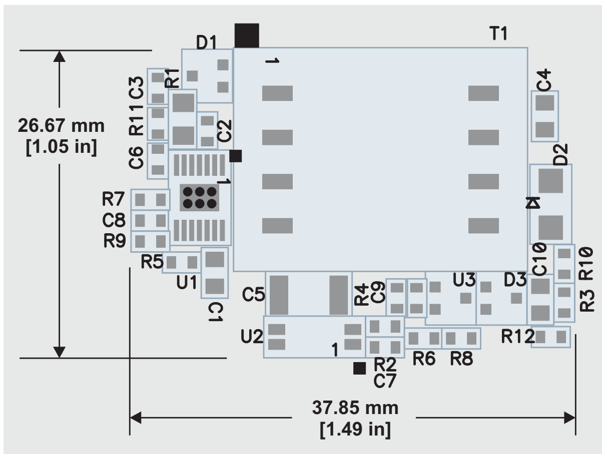
was measured by breaking the loop and injecting a disturbance at R10. In the power stage, the load resistance and output capacitance form a pole at 700 Hz. This pole is compensated for by the zero of the TL431 circuit, formed by the values of C9 and R6. An additional zero was added at 30 kHz by placing C7 in parallel with R2. This zero helps negate the limited bandwidth of the optocoupler and increases the phase margin.

The efficiency of this design is shown in Figure 3. With a typical ON resistance of 130 mΩ, the internal MOSFET contributes around 850 mW of conduction losses with a 5-V input. The output diode dissipates approximately 200 mW. The remaining loss can be attributed to switching losses, bias loss in the TPS61175 and TL431, and losses in the snubber.

This simple design can be placed in a small amount of board space. Figure 4 shows the parts placement for this design on a single-sided PWB. The design consumes a total of 1.5 in^2 on one side of the board.

This simple and compact design demonstrates how integrated boost converters, usually relegated to portable applications, can be leveraged in isolated auxiliary supplies. This example showed how to use the TPS61175 with a low input voltage. With a 40-V rating on the drain of the internal FET, and a maximum input-voltage rating of 18 V, this design could be adjusted to work with a 12-V input, with an internally set peak-current limit of 3 A.

Figure 4. Typical PWB layout



Related Web sites

power.ti.com
www.ti.com/sc/device/TPS61175
www.ti.com/sc/device/TL431

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